Amendments to the Specification:

Please delete the three paragraphs starting on Page 3, line 25 and ending on Page 4, line – and insert the following:

Referring to FIG. 1 of the drawings, a digital communication network 10 includes a radio access network 12 including a base station 14 and a base station controller 16. The radio access network 12 is coupled to a switch fabric 18, which may be a circuit switch network or a packet data network that interconnects the radio access network 12 with a public switched telephone network 22 and other radio or data networks 24. The base station 14 provides wireless communication services to mobile stations 20 operating within a coverage area of the base station 14. Preferably the base station 14 operates in accordance with one or more wireless communication standards, including without limitation a direct sequence code division multiple access (DS-CDMA) system operating in accordance with the IS-~~95~~2000 3G standard.

For any end-user, i.e., mobile station 20, the i[th] chip of an IS-~~95~~2000 3G spread digital signal S can be modeled as:

$$S_i = (P\underline{p}_{[[p]]i} + jDd_iw_i)c_i$$

and consists of a pilot component, $P\underline{p}_{[[p]]i}$, and a data-bearing component[[;]], Dd_iw_i, where P and D are the corresponding amplitudes; p is the pilot sequence; d is the interleaved and possibly-repeated coded information sequence; w is the Walsh-code sequence corresponding to the data-bearing component; and c denotes the product of the short and long pseudo-random noise (PN) sequences.

Please delete the four paragraphs starting on Page 5, line 3 and ending on Page 6, line 24 and insert the following:

To estimate r, ~~*h(0)*~~ $\underline{h^{(0)}}$ and ~~*d(0)*~~ $\underline{d^{(0)}}$ denote the despread pilot component and the despread data component, respectively. An estimate ~~h(1)~~ $\underline{h^{(1)}}$ of Ph is obtained by passing

~~h(0)~~ $\underline{h^{(0)}}$ through a channel estimation filter f; i.e., ~~$hi(1) := (f*H(0))i$~~ $\underline{h_i^{(1)} := (f* h^{(0)})_i}$, where * denotes discrete convolution.

The soft data estimates $d_i^{(1)}$ are obtained as follows: First, the $d_i^{(0)}$, generated by ~~dispreading~~ despreading the data component are phase compensated using the $h_i^{(1)}$,

$$\hat{d}_i := \sum_{a=1}^{A} \sum\nolimits_{m=1}^{M} d_{a,m,i}^{(0)} (h_{a,m,n(i+\delta)}^{(1)})^*$$

where A is the number of receiver antennas; M_a is the number of fingers assigned to resolved rays or multi-path components for antenna a; and x^* denotes the complex conjugate of x. Second, applying a simplifying assumption that ISI_i, TN_i, MAI_i, and the estimation errors in $h_i^{(1)}$ are all uncorrelated and ~~Guassian~~ Gaussian, then

$$\hat{d}_i = j\mu d_i + [[w]]\underline{\varepsilon}_i$$

where $\mu > 0$ and $[[\underline{w}]]\underline{\varepsilon_i}$ denotes a complex-valued, Gaussian random variable whose independent components have mean zero and variance σ^2. Under this assumption, the conditional expectation of d_i given $\hat{d}_i$ is ~~$E[d_i|\hat{d}_i] = \tanh(\mu \mathrm{Im}\{\hat{d}_i\}/\sigma^2)$~~ $\underline{E[d_i|\hat{d}_i] = \tanh(\mu \mathrm{Im}\{\hat{d}_i\}/\sigma^2)}$. Third, μ and σ^2 are estimated on a PCG-by-PCG basis as:

$$[[\sigma^2]]\underline{\hat{\sigma}^2} := 1/(i_2 - i_1)\sum_{i=i_1}^{i_2} (\mathrm{Re}\{\hat{d}_i\})^2$$

$$x = 1/(i_2 - i_1)\sum_{i=i_1}^{i_2} (\mathrm{Im}\{\hat{d}_i\})^2$$

$$\hat{\mu} := \left|x - \hat{\sigma}^2\right|^{1/2}$$

where $i_1 <= i <= i_2$ includes the indices of all coded bits within a specific PCG.

Although the tanh function may be used, in a preferred implementation of the invention, the tanh function is approximated by an applied function t; hence the soft data estimate of d_i is:

$$d_i^{(1)} := t(\hat{\mu}\,\mathrm{Im}\{\hat{d}_i\}/[[\sigma^2]]\underline{\hat{\sigma}^2})$$

A preferred choice of the applied function t is a piece-wise linear function: for $z \in [0,2.4]$, this t is obtained by linear interpolation using the $(z, t(z))$-pairs (0,0), (0.625,0.5721), (1.25,0.8658), and (2.4, 1); for $z > 2.4$, $t(z) := 1$; finally, for $z < 0$, $t(z) := -t(-z)$. The function t is illustrated in FIG. 2.

As will be appreciated from the foregoing discussion, the estimation $\hat{d}_i$ includes an imaginary component and a real component, where the imaginary component is both signal and noise and the real part is only noise. The estimate $[[\sigma^2]]$ $\underline{\hat{\sigma}^2}$ is an estimate of the average noise power while the estimate x is an average of the signal and noise power. Thus, the estimate μ, the difference of x and $[[\sigma^2]]$ $\underline{\hat{\sigma}^2}$, is the signal. It will be further appreciated that the estimation $\hat{d}_i$ is obtained at the chip level, and hence, IC is accomplished at the chip level. A re-spreading operation is performed to generate the "cleaned" signal for the final estimation of the coded information sequence d.

Please delete the four paragraphs starting on Page 7, line 4 and ending on Page 8, line 26 with the following:

For partial interference cancellation, the estimate of ~~r(1)~~ $\underline{r_i^{(1)}}$ of $s_i h_i$ may be written as:

$$r_i^{(1)} := (\alpha_p p_i' + j\alpha_d \eta d_i^{(1)} w_i) c_i h_{n(i)}^{(1)}$$

where α_p and α_d are the partial cancellation coefficients $p_i' = 1$ over the first ¾ of each PCG (i.e. over the known portion of p) and $p_i' = 0$ otherwise; $\eta :=$ D/P; and $d_i^{(1)}$ is an

estimate of d_i. Since the data bits d_i have a higher rate than the output samples of the filter f, the mapping n(·) is needed to match them appropriately: if the sampling rate of f is v_1 Hz and the d_i have a rate of v_2 bits/s, then $n(i) := \cancel{\lfloor iv_1/v_2 \rfloor} \lfloor iv_1/v_2 \rfloor$ (hence, each channel estimate is used for the phase compensation of v_2/v_1 bits).

In accordance with a further preferred embodiment of the invention, the partial interference cancellation coefficients α_p and α_d may also be estimated on a PCG-by-PCG basis. For the purpose of this embodiment, a hard estimate $d_i^{(1)}$ is used and is

$$[[\hat{d}]]\underline{d}^{(1)}{}_i := \mathrm{sgn}(\mathrm{Im}\{\hat{d}_i\})$$

by recalling that the imaginary part of the $\hat{d}_i$ represents only signal, taking the sign of $\hat{d}_i$ is typically used as an estimate. The estimation error of the signal is $(r_i - r_i^{(1)})$, and taking the partial derivative of the estimation error for each of α_p and α_d, respectively, and solving for α_p and α_d provides the following:

$$\alpha_p = \frac{1}{1 + \rho^2 / |Ph(iT_c)|^2}$$

and

$$\alpha_d = \frac{2\beta - 1}{1 + \rho^2 / |Ph(iT_c)|^2}$$

where $\beta := \mathbf{P}[d_i = d_i^{(1)}]$, i.e., the probability that the data estimate is correct and ρ^2 is the variance of the error in estimating the product $Ph(\cdot)$, and wherein T_c is the duration of the chip.

In accordance with the preferred embodiments of the invention, β is determined in real time. Using the simplified statistical model for $\hat{d}_i$ from above, the conditional probability density function of $d_i^{(1)}$ given d_i is Guassian with mean μd_i and variance σ^2. Then, assuming that $\mathbf{P}[d_i = 1] = \mathbf{P}[d_i = -1] = ½$, it follows that

$$1-\beta = P[d_i \neq d_i^{(1)}]$$

$$= (1/\sqrt{\pi}) \int_{-\infty}^{-\mu/\sqrt{2\sigma}} e^{-t^2}\, dt$$

$$= \mathrm{erfc}(\mu/\sqrt{2\sigma})/2$$

where $\mathrm{erfc}(x) := (2/\sqrt{[[2]]\pi}) \int_{x}^{\infty} \exp(-t^2)dt$.

The unknown parameters μ and σ are estimated on a PCG-by-PCG basis as set forth above. Then an estimate of β is:

$$\hat{\beta} := 1 - [[t]]\underline{e}(\hat{\mu}/\sqrt{2\hat{\sigma}})$$

where the approximation $[[t]]\underline{e}(x) \approx \mathrm{erfc}(x)/2$ is introduced for practical implementation. A simple choice for the function $[[t]]\ \underline{e}$ is a piece-wise linear function; for example, for $x \in [0,1.8]$, $[[t]]\ \underline{e}$ is obtained by linear interpolation using the (x, $[[t]]\underline{e}(x)$)-pairs(0, 0.5), (0.8, 0.1), and (1.8, 0); for $x>1.8$, $[[t]]\underline{e}(x) := 0$, as shown in FIG. 4.